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RESEARCH DEPARTMENT

A RADIO-FREQUENCY TRANSFORMER OF CONTINUOUSLY-VARIABLE RATIO (THE 'HELITRAN')

Report No. G-081

(1961/13)

THE BRITISH BROADCASTING CORPORATION
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RESEARCH DEPARTMENT

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SUMMARY

A radio-frequency transformer of continuously-variable ratio has been developed; it has been designed primarily for use in admittance bridges and is particularly useful in bridges for measuring the parameters of transistors.

1. INTRODUCTION

In admittance bridges, the 'unknown' admittance is balanced against a known or 'standard' admittance; the 'standard' usually involves two independently adjustable elements, one corresponding to the in-phase and the other to the phase-quadrature vector components. A type of bridge commonly used is the transformer admittance bridge, shown diagrammatically in Fig. 1.

The unknown admittance y_u, connected to a supply of voltage V, delivers to the detector winding PN a current Vy_u since, when the bridge is balanced, the voltage across PN is zero. The components of the current in the standard are

$-Vg_s$ from the conductance, g_s

and $-jVb_s$ from the capacitance, susceptance b_s.

At balance, $$y_u = g_s + jb_s.$$

In the case of the capacitance, there is no difficulty in providing an adjustment which is continuously variable and capable of being accurately calibrated. The conductance adjustment is less straightforward; the following devices have been used but each presents certain difficulties:

(i) *a wire-wound variable resistance*
This is not continuously variable; moreover, residual inductance and capacitance restrict its use to frequencies below 1 Mc/s.

(ii) *a carbon-track variable resistance*
This has the disadvantage that it is difficult to secure smooth control and lasting accuracy, and also that stray inductance and capacitance are important at frequencies exceeding about 50 Mc/s; moreover, it generally provides a control calibrated linearly in resistance rather than in conductance.

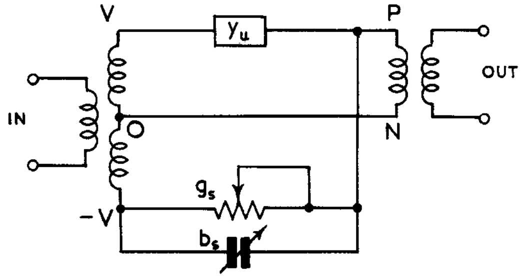

Fig. 1 - Transformer admittance bridge

(iii) *a set of conductances in decade form*

This does not provide continuous variation, requires a large number of precision resistors and is complicated in form; at higher frequencies it is increasingly difficult to maintain the necessary short distance between all effective elements. Nevertheless, decade conductances are used in bridges working up to about 250 Mc/s.

(iv) *a thermistor*

A thermistor can be controlled quite accurately by a servo amplifier over the required range of conductances; it is also adequately compact. Here, the difficulty is to provide isolation between the radio-frequency circuits and the thermistor control circuits without imposing limitations affecting the accuracy. In particular, it is difficult to organize a thermistor bridge for three-terminal measurements, i.e. with the connexion N in Fig. 1 accessible and usable. In brief, a thermistor must have series capacitances in its radio-frequency circuits and series inductances in its control circuits in order to prevent mutual interference.

There is, however, another method of obtaining the effect of a variable conductance and this is described in Section 2.

2. PREFERRED FORM OF VARIABLE CONDUCTANCE

Fig. 2 shows how a potentiometer resistance R, in association with a fixed conductance g_o, can be used to act as a variable conductance. If the potentiometer setting is λR from the zero voltage position and $(1-\lambda)R$ from the maximum voltage position, it behaves as a source of e.m.f. λV. Its contribution to the detector current is, however, not proportional to λ but is

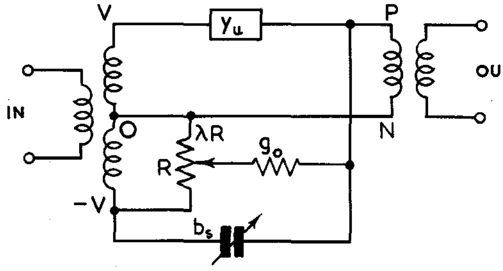

Fig. 2 - Bridge incorporating a resistance potentiometer

$$i_g = \lambda V . \frac{g_o}{1+\lambda(1-\lambda)Rg_o}$$

owing to the variable resistance $[\lambda.R$ in parallel with $(1-\lambda)R]$ in series with the standard g_o. Thus if R were 100 ohms and g_o were 1 mmho:

$$\frac{i_g}{V} = \frac{\lambda}{1+\frac{\lambda(1-\lambda)}{10}} \text{ mmho}$$

Thus the effective conductance is λ mmho with a fractional error, in this example, of $\lambda(1-\lambda)/10$; this is maximum when $\lambda = \frac{1}{2}$ and is then 2·5%.

The arrangement of Fig. 2 has many advantages; these include the property that zero effective conductance can be obtained with finite values of R and g_o, and that the device can be approximately linear in conductance. More important is that much higher residual inductances or capacitances can be tolerated in the potentiometer than in a variable resistive standard, since R can be considerably smaller than the resistance r_o $(=1/g_o)$. Thus, under certain circumstances, the resistive potentiometer could be replaced by an equivalent arrangement of two inductors or two capacitors, as in Fig. 3. Regarded as a generator, the potentiometer will now present a source impedance consisting of C_1 and C_2 in parallel. This capacitive reactance causes a frequency-dependent quadrature error which cannot be allowed for by calibration, as may be done in the case of the resistive potentiometer. The capacitors C_1 and C_2 must therefore be made large enough at the lowest operating frequency for this error to be small. This results, unfortunately, in a low impedance shunting the supply which seriously reduces the sensitivity at high frequencies.

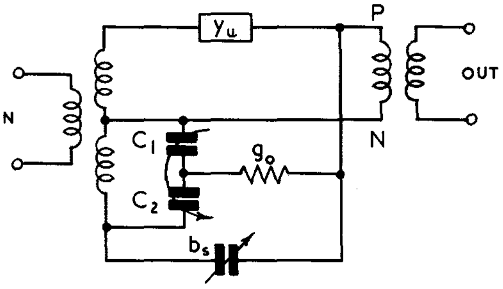

Fig. 3 - Bridge incorporating a reactance potentiometer

If, instead of a potentiometer (whether resistive or reactive), mutual inductance is used (i.e. a variable-ratio transformer), enormous advantages accrue. The series impedance imposed by the device is now the short-circuit reactance,* and the load imposed on the supply (equivalent to R with the potentiometer in Fig. 2) is now the open-circuit reactance.** With the potentiometer, the ratio between the load on the supply and the maximum additional series impedance is four; with the variable transformer or autotransformer, ratios of the order of hundreds may be obtained. For example, an autotransformer used at 1 Mc/s may have an open-circuit reactance of, say, 60 ohms, so that it does not unduly load the supply; the short-circuit reactance may be made as low as 0·2 ohms, and this would be quite negligible in comparison with $r_o (=1/g_o) = 1000$ ohms. The situation is not quite so favourable as this, since a bridge with a lower frequency limit of 1 Mc/s would probably be required to work at frequencies up to about 30 Mc/s; the open-circuit reactance would then be unnecessarily high and the series error 6 ohms instead of 0·2 ohms. The same considerations apply to devices designed to operate up to 300 Mc/s, namely that the open-circuit impedance of the device must be high enough for satisfactory sensitivity at the lowest operating frequency, and the short-circuit reactance must be such as to make measuring errors small at the highest operating frequency. Thus the limiting feature in practice is the range of frequencies over which the device is required to operate.

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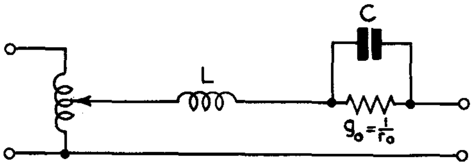
*i.e. the reactance effectively in series with the tapping point when the input winding is short-circuited.

**i.e. the reactance of the input winding with the tapping point open-circuited.

The error caused by the reactance in series with the standard conductance is principally a phase error in the current flowing through the conductance to the detector; it will therefore appear mainly as a susceptance error when the device is used in an admittance bridge. If the short-circuit reactance of the transformer is expressed as mr_o, where $r_o = 1/g_o$ and m ($\ll 1$) varies with the frequency, the phase error will be m radians and the fractional amplitude error will be $\frac{1}{2}m^2$. Thus, to keep errors within 1%, the maximum usable frequency is that for which $m = 0{\cdot}01$. However, the phase error can be greatly reduced by a simple addition to the circuit as is shown in Fig. 4. Here, a compensating capacitance C with a susceptance of mg_o is connected in parallel with the conductance g_o. The reactance of the short-circuit inductance L is expressed as $(m + \delta)r_o$ where δ represents a small variation depending on the transformer setting. Using this compensated circuit, it can be shown that the phase error becomes $(m^3 + \delta)$ radians and that the fractional amplitude error becomes m^2. If $\delta \ll m$, the compensation allows m to rise to $0{\cdot}1$ before the error exceeds 1%, that is, ten times the frequency usable without compensation. It is found, however, that in the present v.h.f. helitran, the variation δ is about $\pm\, 0{\cdot}3\, m$ and this represents the limiting feature.



$\omega L = (m + \delta)\, r_o$
$\omega C = mg_o$

Fig. 4 - Compensation for mean short-circuit reactance of a transformer

3. THE 'HELITRAN'

We now consider the design of the autotransformer to satisfy the requirements outlined above. To this autotransformer we have given the name HELITRAN.

It is basically a simple helical conductor with a concentric shaft carrying a contact which may be adjusted to connect to any part of any turn from start to finish. The mechanics of this helitran are similar to those of the well-known 'helipot' which is a resistive potentiometer of several turns; in the helitran, the resistive element is replaced by a good conductor.

The errors will be divided into two categories, namely

(a) those due to imperfect coupling between turn and turn
and (b) those due to imperfect coupling of a submultiple of a turn to the whole.

3.1. Imperfect Coupling between Turn and Turn

Imperfect turn-to-turn coupling has two important effects:

(i) it reduces the open-circuit reactance and so reduces sensitivity
and (ii) it increases the short-circuit reactance, and hence the errors.

With a plain helix, these effects would generally be excessive; there would also exist a concomitant error that the voltage per turn, even at zero load, would vary from turn to turn. Several methods are available to improve this situation. The most obvious is to make the effective axial distance between turns as small as possible. For this purpose the conductor should have a basically rectangular cross-section, as shown in Fig. 5, with the insulation between turns as small as convenient. This means alone, however, is not sufficiently effective with feasible conductor and insulation dimensions.

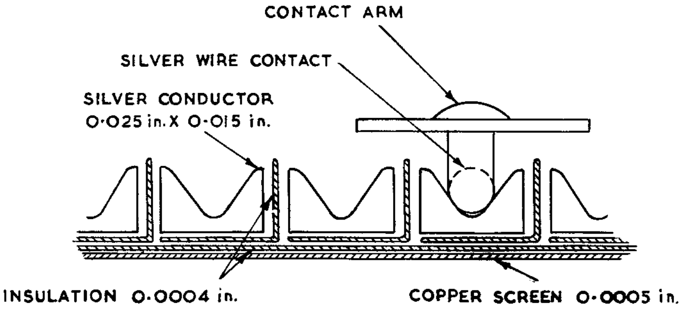


Fig. 5 - Cross-section through winding of h.f. helitran

The coupling can be improved by using a screen of copper foil with an insulated overlap parallel to the axis, so that each turn links all other turns by linking the screen. This gives satisfactory results at frequencies up to about 30 Mc/s. The lower frequency limit can be extended to below 1 Mc/s by using a ferrite core inside the copper screen.

A third method is to connect each turn of the helix to the corresponding turn on a special compensating transformer, so that the leakage between turns is that between the compensating transformer turns, which can be made very small. This construction results in a very low short-circuit inductance and makes the device useful at frequencies as high as some hundreds of megacycles per second. In certain cases it may not be necessary to connect every turn on the compensating transformer to the corresponding turn on the helitran; it may be sufficient to connect only every second turn, or even every third turn, thus simplifying the construction.

In some applications it may be advantageous to connect the common terminal of the supply to one of the intermediate transformer tappings so that, as the slider crosses this point, the current through the standard impedance falls to zero and rises in the opposite phase. This is useful in transistor-measurement bridges where admittances may be in any part of the complex plane, and in special bridges for measuring reflexion coefficients in transmission-line work.

3.2. Imperfect Coupling of a Submultiple of a Turn to the Whole

Although the device is basically perfect as a voltage divider when open-circuited, and as a current divider when short-circuited, an error occurs when the load is finite; this is caused by poor coupling between the parts of a turn on each side of the sliding contact and appears as a leakage inductance in series with the load. The leakage is least when the contact arm crosses the position of the radial lead-out from the shaft, and is greatest when the contact arm is in the opposite position. This variation of leakage inductance as the setting of the contact is changed gives rise to errors which are not easily compensated, as discussed in Section 2. One way of reducing the variation, for a given open-circuit inductance, is to make the area of each turn smaller and, at the same time, increase the number of turns. In practice, the use of ten turns gives a reasonable compromise between the errors discussed above and errors due to imperfect turn-to-turn coupling discussed in Section 3.1. It is also convenient mechanically since index mechanisms for ten revolutions of a shaft are readily available.

The connexion of the sliding contact to the output terminal viâ the rotating shaft introduces additional leakage which, although constant, should preferably be made as small as possible. Where the contact is made inside the coil, the rotating shaft is therefore fitted with a thin radial 'paddle' (Fig. 6); this increases the path length of the associated magnetic flux without impeding the axial magnetic flux, since the paddle has a small area projected in the axial direction. Both the paddle and the lead-out from the shaft are made thin in section so that the flux within the helix is kept relatively uniform. If, for example, the fixed lead-out were thick, then, as the half of the paddle carrying the contact passes close to the lead-out, it would traverse a region of almost zero flux. The output would then tend to remain constant over a finite sector instead of continuing to vary linearly with the rotation of the paddle.

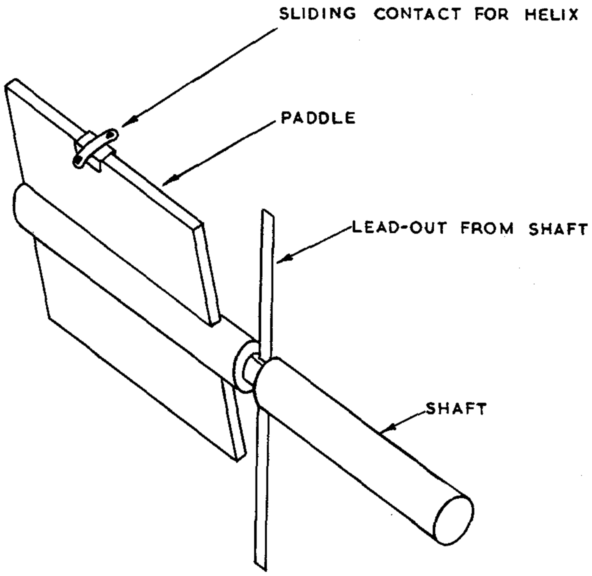


Fig. 6 - Radial paddle of v.h.f. helitran, showing contacts

Another method of reducing this error has been suggested, but not yet verified. It consists of placing two D-shaped loops of wire on the paddle, connected together as a figure of eight. This will have the effect of constraining the axial flux on each side of the paddle to be equal and in the same sense, and will give a minimum error at points half-way between the existing minima.

3.3. Examples of Practical Designs

By way of example, two helitran designs are described below and shown in Fig. 7. One is a h.f. helitran, designed for use in an admittance bridge working at

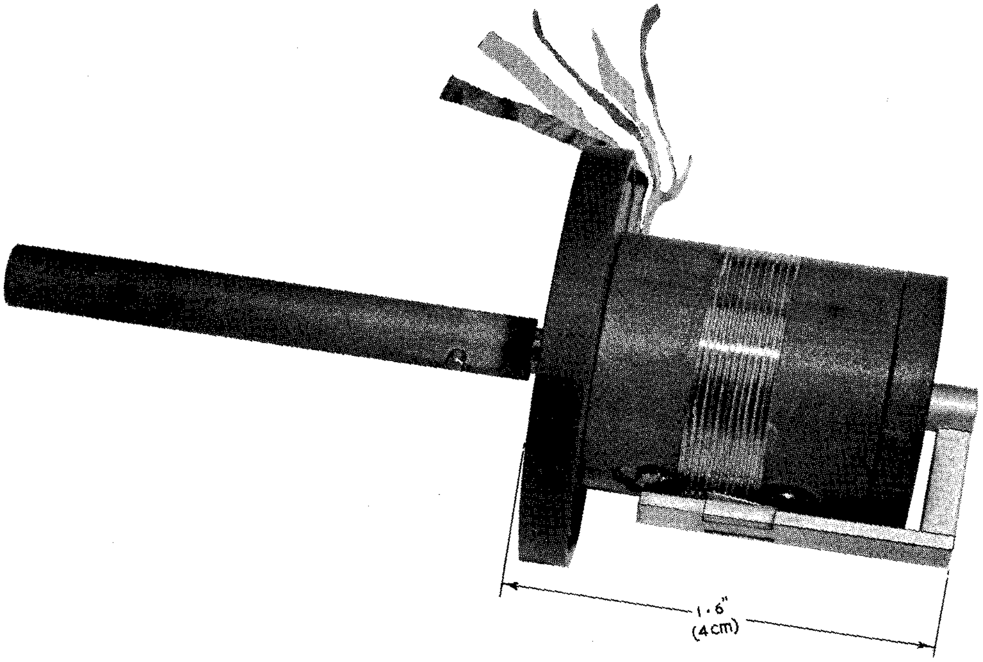


Fig. 7(a) - H.F. helitran

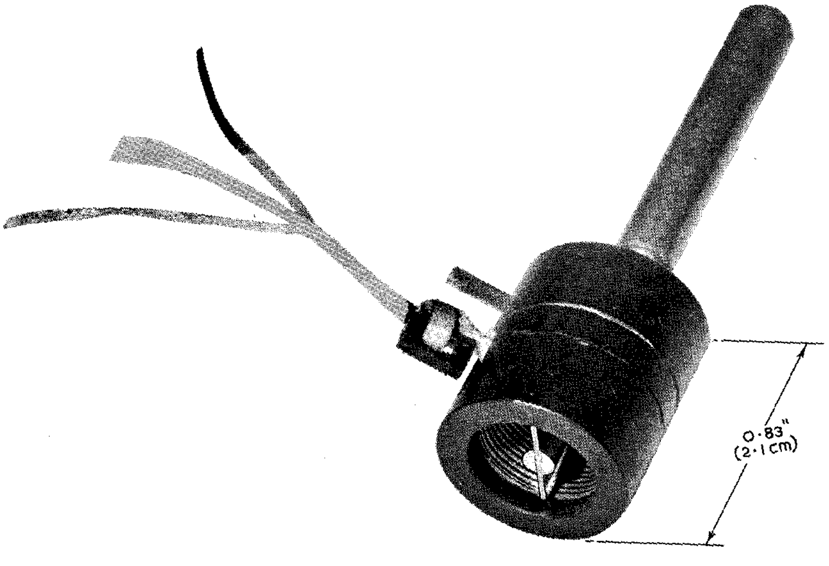


Fig. 7(b) - V.H.F. helitran

frequencies between 1 and 30 Mc/s, and the other is a v.h.f. helitran for a similar bridge working at frequencies between 30 and 300 Mc/s.

The significant features to be tested are as follows:

(i) *Open-circuit input impedance.* This fixes the loading imposed by the device on the generator supplying the bridge. If the impedance is too low, the sensitivity may be limited unless a suitable transformer is employed to improve the impedance match.

(ii) *Reactance measured at variable tap with input short-circuited.* This will generally be considered as having a mean value, a variation from turn to turn and an additional variation over each turn. In general the mean reactance can be compensated but not the variable part.

(iii) *The voltage change per unit change of index angle.* This should ideally be constant over the whole range of variation.

3.3.1. H.F. Helitran (1–30 Mc/s)

The h.f. model has a sliding contact which is outside the helical winding, as shown in Fig. 7(a). The coil has ten turns with a diameter of 1 in. (2·5 cm), giving an open-circuit inductance of about 4 μH, but the figure is increased to about 8 μH in this model by using a cylindrical ferrite core. The open-circuit reactance is therefore satisfactory at frequencies down to below 1 Mc/s when working from a source of about 60 ohms impedance.

Regarding the short-circuit reactance, the complication of a compensating transformer is not warranted at these relatively low frequencies. Instead, the helix is wound over a copper screen with insulation 0·001 in. (0·025 mm) thick. The sliding contact is carried on an arm running close to the outside of the helix. The short-circuit inductance, shown in Fig.8, is approximately 56 mμH ± 35 mμH, the variation corresponding to a reactance of ±7 ohms at 30 Mc/s. If the load is 1000 ohms, the variable error (which is in quadrature) is therefore ± 0·7% at 30 Mc/s.

The deviation from linearity was too small to be detected with available apparatus; this means it did not exceed 0·2%.

3.3.2. V.H.F. Helitran (30–300 Mc/s)

The v.h.f. model has a sliding contact which is inside the helical winding, as shown in Fig.7(b). The ten-turn coil has a diameter of ½ in. (1·3 cm) and an open-circuit inductance of about 1 μH, which is satisfactory down to a frequency of 10 Mc/s. In this case a compensating transformer (see Section 3.1) has been used to reduce the short-circuit reactance to a satisfactory value. The transformer itself consists of ten turns of insulated copper tape on a miniature ferrite core; the transformer leads are laid in a slot 0·1 in. (0·25 cm) wide running axially along the tube containing the helix. Connexions are made to the outside of the helix, the sliding contact being carried on the internal paddle.

The mean short-circuit inductance, shown in Fig.9, is 21 mμH; the variation of ± 6 mμH corresponds to a variable reactance of ± 3·6 ohms at 100 Mc/s. If the load is 1000 ohms the variable error (which is in quadrature) is ± 0·36%.

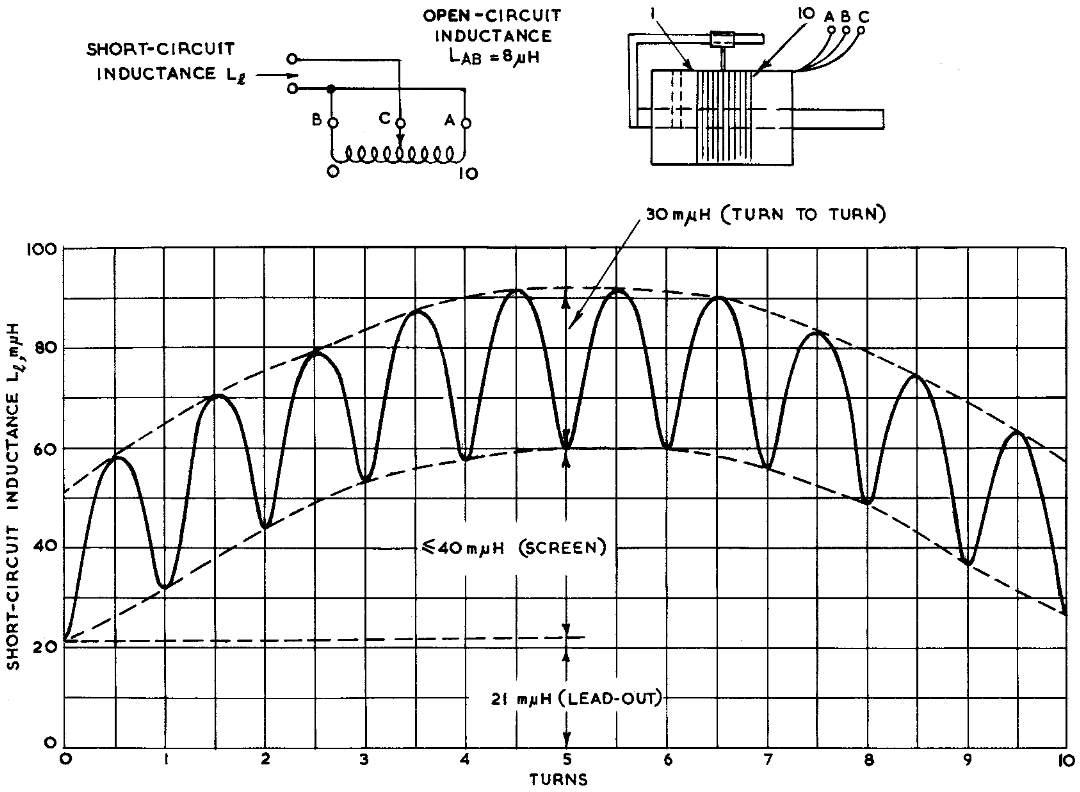


Fig. 8 - Short-circuit inductance of h.f. helitran (1-30 Mc/s)

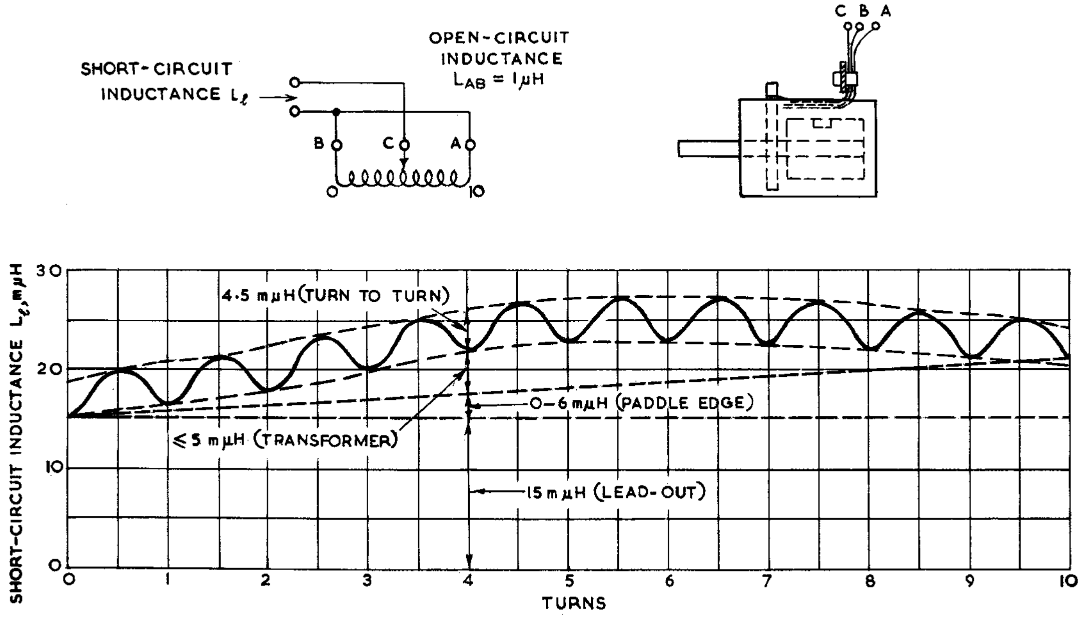


Fig. 9 - Short-circuit inductance of v.h.f. helitran (30-300 Mc/s)

The v.h.f. helitran can be used in a bridge at much lower frequencies, say at 1 Mc/s, by feeding it from a suitable tapping on the bridge ratio-arm transformer. If a 5/1 turns ratio is used, this would raise the input impedance from 6 ohms to 150 ohms. The short-circuit reactance of the helitran is only 36 mmho at this frequency. Consequently, although the voltage applied to it is now reduced to a fifth of the bridge input voltage, the maximum current (and therefore the effective conductance range) can be maintained without increasing the errors by using a higher value of standard conductance, say 5 mmho (r_0 = 200 ohms) instead of 1 mmho (r_0 = 1000 ohms). In essence, the helitran must be matched in and out at the operating frequency to secure best results.

4. CONCLUSIONS

The two designs of helitran described promise to be very useful components for accurate and compact admittance bridges. In particular, used as a precision voltage divider, a helitran can simplify the design of the variable conductance control of three-terminal bridges which employ a transformer ratio-arm; bridges for measuring r.f. transistor parameters represent one field of application.

Recent work has suggested that, where additional fixed tappings are not required, the construction of the v.h.f. helitran could be greatly simplified. It has been found that adequate inter-turn coupling can be provided by a single-turn outer screen which acts in the same way as the inner screen of the h.f. helitran but has thinner insulation. The multi-tap transformer is not then required to keep the leakage reactance at a low value.

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